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كلية الدراسات العليا

التاريـــخ: ١٧ / ٩ / ٢٠١١

نموذج رقم (١٨)
اقرار والتزام بقوانين الجامعة الأردنية وأنظمتها وتعليماتها لطلبة الماجستير

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User location and Mobility effect on Orthogonal frequency Division Multiple Access Systems' Capacity.

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USER LOCATION AND MOBILITY EFFECT ON ORTHOGONAL FREQUENCY DIVISION MULTIPLE ACCESS SYSTEMS CAPACITY

By

Suhad Atallah Abu-Elzait

Supervisor

Dr. Jamal S. Rahhal.

This Thesis was Submitted in Partial Fulfillment of the Requirements for the Master's Degree of Science in Communications Engineering

Faculty of Graduate Studies

The University of Jordan

September, 2011

COMMITTEE DECISION

This Thesis/Dissertation (User Location And Mobility Effect On Orthogonal Frequency Division Multiple Access Systems Capacity) was Successfully Defended and Approved on ----٢٠١١/٨/١١----

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Dedicated to my parents for their love, understanding, and support

and my two little angels, Celine and Masah, with all my love…

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LIST OF ABBREVIATIONS

AMC adaptive modulation and coding
AoA angle of arrival
AWGN additive white Gaussian noise
BEP bit error probability
BER bit error rate
BS base station
CDF cumulative distribution function
CDMA code division multiple access
CP cyclic prefix
CSI channel state information
DFT discrete Fourier transform
DL downlink
DS delay spread
DVB-H digital video broadcasting-handheld
FDMA frequency division multiple access
FEQ frequency-domain equalization
FFT fast Fourier transform
FIR finite impulse response
HDTV high-definition television
HIPERMAN high-performance metropolitan area network
HO handover
ICI intercarrier interference
IFFT inverse fast Fourier transform
LOS line of sight
LTE long-term evolution
MIMO multiple input multiple output
MS mobile station
OFDM orthogonal frequency division multiplexing
OFDMA orthogonal frequency division multiple access
PDF probability density function
P/S parallel to serial
QoS quality of service
QAM quadrature amplitude modulation
SINR signal-to-interference-plus-noise ratio
SNR signal-to-noise ratio
S/P serial to parallel
STBC space/time block code
VoD video on demand
Wi-Fi wireless fidelity
WiMAX worldwide interoperability for microwave access

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ABSTRACT

One advantage of using OFDM/OFDMA compared to other techniques is obtained when the channel is highly frequency selective. For single carrier techniques, a highly frequency selective channel usually implies that a very complex channel equalizer is needed in order to get reliable communications. For a system based on OFDM, which is a multicarrier system, this is avoided by transmitting the information on many orthogonal sub-carriers. In a multicarrier system the symbol rate on each sub-carrier can be lower and the resulting longer symbol duration makes it easier to offer higher immunity against multipath dispersion.

Critical for an OFDM system to work properly is that the orthogonality between the sub-carriers is maintained. If this is not achieved, then the performance is degraded considerably. There are different reasons why orthogonality between the sub-carriers might be lost, e.g., frequency error, time-drift, phase noise, and perhaps most importantly Doppler spread. Doppler spread is experienced when the receiver is moving and the received signal consists of several paths, each arriving at different angles. The problem with the Doppler effect is mainly due to inter-carrier interference (ICI). This means that information from one sub-carrier leaks over to the neighbouring sub-carriers.

This study tried to reduce the impact of Doppler effect on the mentioned system performance. A mobile unit during motion, is subject to a handover to a neighbouring base station. Traditionally, the point of decision is power level, but according to this search, another fact is taken into consideration. That factor is the level of Doppler interference, which might vary this decision to the fever of another base station. This might reduce the rate assigned to such a unit and/or power level, but with a good enhancement in the quality of the signal. According to the obtained results, a maximum of 3٥% of capacity enhancement was achieved at higher signal-to-noise-ratios. Meanwhile an enhancement of 5% was obtained at lower signal-to-noise-ratios.

CHAPTER ONE

MIMO-OFDMA SYSTEMS

Introduction

Orthogonal Frequency-Division Multiple Access (OFDMA) is a multi-user version of the popular Orthogonal frequency-division multiplexing (OFDM) “digital modulation scheme”.

It can be seen as an alternative to combining OFDM with time division multiple access (TDMA) or time-domain statistical multiplexing, i.e. packet mode communication. Where low-data-rate users can send continuously with low transmission power instead of using a "pulsed" high-power carrier, constant delay, and shorter delay, can be achieved.

OFDMA is considered as highly suitable for broadband wireless networks, due to advantages including scalability and MIMO-friendliness, and ability to take advantage of channel frequency selectivity.[1], in addition to its allowance to utilize the capacity features of MIMO systems.

Hence a general outlook at the main features of those two systems, pointing to OFDM systems and MIMO systems, becomes a must ever since they represent the basis of our selected system of research.

(1) Orthogonal Frequency Division Multiplexing (OFDM)

Orthogonal frequency division multiplexing (OFDM) is a multicarrier modulation technique that has recently found wide adoption in a variety of high-data-rate communication systems, including digital subscriber lines, wireless LANs (802.11a/g/n), digital video broadcasting, and now WiMAX and other emerging

wireless broadband systems such as the proprietary Flash-OFDM developed by Flarion (now QUALCOMM), and 3G LTE and fourth generation cellular systems.

Its popularity for high-data-rate applications stems mainly from its efficient and flexible management of intersymbol interference (ISI) in highly dispersive channels.

If T_s indicates symbol duration (time) and τ indicates the delay spread of the channel, by definition, a high-data –rate system will generally have $\tau \rangle\rangle T_s$ since the number of sent symbols per second is high, and accordingly T_s is small. And as a result, all types of wireless broadband systems will suffer from severe ISI and hence will require transmitter and/or receiver techniques that overcome the ISI.

Multicarrier Modulation

The competing desire for high data rates and ISI-free channels, led to what is called “Multicarrier modulation”, since a digital communication system cannot function if ISI is present.

Multicarrier modulation splits the high-rate transmit bit stream into multiple L lower-rate substreams. Each substream has $T_s / L \rangle\rangle \tau$ and $T = LT_s$, where T is the new symbol time. These individual substreams are then sent over parallel subchannels with the data rate on each of these subchannels much less than the total data rate, and so maintaining the total desired data rate, but with less bandwidth. Hence the subchannel bandwidth is much less than the total system bandwidth.

The number of substreams "L" is chosen to guarantee that each subchannel has a bandwidth less than the coherence bandwidth of the channel, so that the subchannel appears as a flat-fading channel instead of a frequency-selective channel, thus the ISI on each subchannel is small.

Each smaller data stream is then mapped to individual data sub-carrier and modulated using some sorts of PSK (Phase Shift Keying) or QAM (Quadrate Amplitude Modulation). i.e. BPSK, QPSK, 16-QAM, 64-QAM.

In FDM, the carriers are nonorthogonal, and so are largely spaced, meanwhile typically, the are orthogonal in OFDM and hence called Orthogonal FDM. The concept of orthogonality is hidden under the cover of Fourier transform theory.

OFDM Basics

Multicarrier modulation requires L independent RF units and demodulation paths, which is a serious drawback, in addition to the need of a very high quality and thus expensive LPF to maintain the orthogonality of the subcarriers at the receiver. So to overcome these needs, OFDM uses the discrete Fourier transform computational technique (DFT) which is highly efficient implemented by the (FFT) algorithm. This low complexity algorithm (FFT) and its inverse (IFFT), and by using a single radio can create a multitude of orthogonal subcarriers.

The channel spread is one of the important channel information, is given for a finite length channel as,

$$H(z) = \sum_{k=0}^{L} h[k] z^{-k} \tag{1.1}$$

Where the channel order L is known at the transmitter while the amplitudes h[k] are unknown. The frequency response of the FIR channel is represented as:

$$H(e^{j2\pi fT}) = \sum_{k=0}^{L} h[k] e^{j2\pi kT} \tag{1.2}$$

Where T is the symbol duration or the sampling period. Since $H(e^{j2\pi fT})$ is the frequency response of the channel h[k]=h(kT), hence, it is a periodic function of f with period $\frac{1}{T}$.

The DFT which is a sampled function of the channel frequency response is defined by

$$DFT\{x[n]\} = X[m] \triangleq \frac{1}{\sqrt{L}} \sum_{n=0}^{L-1} x[n] e^{-j\frac{2\pi nm}{L}} \tag{1.3a}$$

and its inverse is given by

$$IDFT\{X[m]\} = x[n] \triangleq \frac{1}{\sqrt{L}} \sum_{m=0}^{L-1} X[m] e^{j\frac{2\pi nm}{L}} \tag{1.3b}$$

for a point L.

If N is the total number of uniform samples in each frequency period $\frac{1}{T}$, then

$$\begin{aligned} f_0 &= 0.\frac{1}{NT} = 0 \\ f_1 &= 1.\frac{1}{NT} = \frac{1}{NT} \\ &\vdots \\ f_{N-1} &= (N-1).\frac{1}{NT} = \frac{(N-1)}{NT} \end{aligned} \tag{1.4}$$

A simple notation can be used to describe the DFT sequencer by substituting

$$w_n = \frac{2\pi n}{NT}$$

$$\begin{aligned} H[n] &= H(e^{j2\pi fT}) \\ &= \sum_{k=0}^{L} h[k] \exp(-jw_n Tk) \\ &= \sum_{k=0}^{L} h[k] \exp\left(-j2\pi \frac{n}{NT} kT\right) \\ &= \sum_{k=0}^{L} h[k] \exp\left(-j2\pi \frac{nk}{N}\right) \quad n = 0,1,..,(N-1) \end{aligned} \tag{1.5}$$

Notice that H[n] is periodic in N. Therefore H [-n] = H[N-n],

The DFT is highly implemented with the Fast Fourier transform (FFT) algorithm which is of low complexity in turn, and used to make the OFDM realizable in practice.

By applying the DFT theory the orthogonality of the subcarriers is satisfied, now in order to gain orthogonality between OFDM symbols -which consist of L symbols grouped together – a guard band "Tg" must be added between subsequent symbols satisfying the condition that Tg $>\tau$. i.e. a time guard band which is greater than the delay spread of the channel. As long s this condition is satisfied each OFDM Symbol will interfere only with itself, and the next step is to remove the ISI within each OFDM symbol.

The solution to this situation appears also in the FFT algorithm, but this time with a small enhancement. In order for the IFFT/FFT to create an ISI-free channel the channel must appear to provide a circular convolution. As we know when a signal is transmitted through a linear channel nature provides a linear convolution, between the transmitted signal and the channel response.

$$y[n] = x[n]*h[n] \tag{1.6}$$

Where y[n] is the output of the linear time – invariant (FIR) channel, and X[n] is the input of the channel and h[n] is the channel response.

According to the following linear convolution relationship

$$z[k] = \sum_{k=0}^{L} h[i]ss[k-1] + w[k] \tag{1.7}$$

A vector of N output symbols can be written in matrix form as:

$$\begin{pmatrix} z[N] \\ z[N-1] \\ \vdots \\ z[L] \\ \vdots \\ z[1] \end{pmatrix} = \begin{pmatrix} h[0] & h[1] & \ldots & h[L] & & & & \\ & & & \ldots h[L] & & & & \\ & \ddots & & \ddots & & \ddots & & \\ & & & h[0] & h[1] & \ldots & h[1] & \\ & & & \ddots & \ddots & \ddots & \ddots & \\ & & & & & h[0] & h[1] \ldots & h[L] \end{pmatrix}$$

$$\times \begin{pmatrix} ss[N] \\ ss[n-1] \\ \vdots \\ ss[1] \\ ss[0] \\ \vdots \\ ss[-(L-1)] \end{pmatrix} + \begin{pmatrix} w[N] \\ w[N-1] \\ \vdots \\ w[L] \\ \vdots \\ w[1] \end{pmatrix} \tag{1.8}$$

Now to compute the system output in terms of a circular convolution we will have:

$$y[n] = x[n] \otimes h[n] = h[n] \otimes x[n] \overset{\Delta}{=} \sum_{k=0}^{L-1} h[k] x[n-k]_L \tag{1.9}$$

Where the circular function x[n] = x[nmodL] is a periodic version of $x[n]$ with period L.

If we take the (DFT) of the channel output y[n]:

$$DFT\{y[n]\} = DFT\{h[n] \otimes x[n]\} \tag{1.10}$$

That gives the frequency domain

$$Y[m] = H[m]X[m] \tag{1.11}$$

This duality between the circular convolution in time domain and simple multiplication in frequency domain is a property unique to the DFT.

So if the channel – frequency response is known to the receiver it is trivial to recover the input symbol simply by computing

$$\hat{X}[m] = \frac{Y[m]}{H[m]} \tag{1.12}$$

Where $\hat{X}[m]$ is the estimated value of X[m], which in turn is imperfect due to additive noise, cochannel interference, imperfect channel estimation, nevertheless, the ISI, which is the most serious form of interference in a wideband channel, has been mitigated.

But as mention earlier, that is not what really occurs in reality, so in order to achieve such an advantages circular convolution is faked, that happens by adding a specific prefix called cyclic prefix (CP) no to the transmitted vector. This creates a signal that appears to be x[n]L and so y[n]=[x]*h[n].

If the minimum channel delay spread has duration of (v +1) samples then adding a guard band of at least (v) samples, between OFDM symbols makes subsequent symbols independent of each other.

For a single OFDM symbol:

$$\begin{aligned} X &= [x_1 \ x_2 \ldots x_L] \\ or \ &= [x_0 \ x_1 \ldots x_{L-1}] \end{aligned} \tag{1.13}$$

Adding a cyclic prefix of length (v) gives:

$$X_{cp} = [\underbrace{x_{L-v} \ x_{L-v+1} \ldots x_{L-1}}_{\text{Cyclic Prefix}} \ \underbrace{x_0 \ x_1 \ldots x_{L-1}}_{\text{Original Date}}] \tag{1.14}$$

The channel output then is by definition, the linear convolution $\overrightarrow{y_{cp}} = \vec{h} * \overrightarrow{X}_{cp}$, with $\vec{h}$ as a vector of length (v+1), and $\overrightarrow{X}_{cp}$ of length (L+v) as stated earlier, then the resulting $\vec{y}_{cp}$ will be a vector of length (L+2v) samples. Where the first (v) symbols which contains the interference from the preceding OFDM symbol are discarded, as well as the

last v symbols which disperse into the subsequent OFDM symbol, leaving exactly "L" samples for the desired recovery of $\vec{y}$.

The most important fact here is that the "L" samples of $\vec{y}$ are equivalent to $\vec{y} = \vec{h} * \vec{x}$.[2,5,9].

So by mimicking a circular convolution, a cyclic prefix that is at least as long as the channel duration allows the channel, output $\vec{y}$ to be decomposed into a simple multiplication of the channel frequency domain input and the channel frequency response.

A detailed analysis is as follows:

To introduce the cyclic prefix in the transmitted data. This step replaces the M leading elements.

ss[0], ss[1], …, ss[-(L-1)]

Of the (N+L) dimensional data vector by the trailing symbols

{ss[N], ss[N-1], …, ss[N-L+1] $\rightarrow$ {ss[0], ss[1], …., ss[-(L-1)]}

By inserting the cyclic prefix we can write the matrix as:

$$\begin{pmatrix} z[N] \\ z[N-1] \\ \vdots \\ z[L] \\ \vdots \\ z[1] \end{pmatrix} = \begin{pmatrix} h[0] & h[1] & \ldots & h[L] & & & & & & \\ & & & \ldots & h[L] & & & & & \\ & & & \ddots & \ddots & \ddots & & & & \\ & & & & & h[0] & h[1] & \ldots & h[L] & \\ & & & & & \ddots & \ddots & \ddots & \ddots & \\ & & & & & & h[0] & h[1] & \ldots & h[L] \end{pmatrix}$$

$$\times\begin{pmatrix} ss[N] \\ ss[n-1] \\ \vdots \\ ss[1] \\ ss[N] \\ \vdots \\ ss[N-(L-1)] \end{pmatrix} + \begin{pmatrix} w[N] \\ w[N-1] \\ \vdots \\ w[L] \\ \vdots \\ w[1] \end{pmatrix} \tag{1.15}$$

$$= \begin{pmatrix} h[0] & h[1] & ... & h[L] & 0 & ... & 0 \\ 0 & h[0] & h[1] & ... & h[L] & \ddots & \vdots \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \ddots \\ 0 & \ddots & 0 & h[0] & h[1] & \dots & h[L] \\ h[L] & \ddots & \ddots & 0 & h[0] & \dots & h[L-1] \\ h[1] & \dots h[L] & 0 & \dots & 0 & & h[0] \end{pmatrix}$$

$$\times\begin{pmatrix} ss[N] \\ ss[n-1] \\ \vdots \\ ss[1] \end{pmatrix} + \begin{pmatrix} w[N] \\ w[N-1] \\ \vdots \\ w[L] \\ \vdots \\ w[1] \end{pmatrix} \tag{1.16}$$

The critical role in the cyclic prefix is to convert the convolutional channel matrix into a well structure N×N cyclic matrix H_{cp}.

The N point DFT can be written in matrix form as;

$$W_N \stackrel{\Delta}{=} \begin{pmatrix} 1 & 1 & \cdots & 1 \\ 1 & W_N^1 & \cdots & W_N^{(N-1)} \\ \vdots & \vdots & \cdots & \vdots \\ 1 & W_N^{(N-1)} & \cdots & W_N^{(N-1)^2} \end{pmatrix} \tag{1.17}$$

And the N point IDFT as:

$$W_N^{-1} = \frac{1}{N}\begin{pmatrix} 1 & 1 & \cdots & 1 \\ 1 & W_N^{-1} & \cdots & W_N^{-(N-1)} \\ \vdots & \vdots & \cdots & \vdots \\ 1 & W_N^{-(N-1)} & \cdots & W_N^{-(N-1)^2} \end{pmatrix} \tag{1.18}$$

An amazing property of the cyclic prefix can be established by applying the DFT and IDFT matrices.

$$H_{cp}.W_N^{-1} = \frac{1}{N}\begin{pmatrix} H[0] & H[-1] & \cdots & H[-N+1] \\ H[0] & H[-1]W_N^{-1} & \cdots & H[-N+1]W_N^{-(N-1)} \\ \vdots & \vdots & \cdots & \vdots \\ H[0] & H[-1]W_N^{(N-1)} & \cdots & H[-N+1]W_N^{-(N-1)^2} \end{pmatrix}$$

$$= W_N^{-1}\begin{pmatrix} H[0] & & & \\ & H[-1] & & \\ & & \ddots & \\ & & & H[-N+1] \end{pmatrix}$$

$$H_{cp}.W_N^{-1} = W_N^{-1}.D_H \tag{1.19}$$

From the periodic nature of H[n]

$$D_H = \begin{pmatrix} H[0] & & & \\ & H[-1] & & \\ & & \ddots & \\ & & & H[-N+1] \end{pmatrix} = \begin{pmatrix} H[N] & & & \\ & H[-1] & & \\ & & \ddots & \\ & & & H[1] \end{pmatrix} \tag{1.20}$$

$$H_{cp} = W_N^{-1} H_{cp} W_N \tag{1.21}$$

$$\begin{pmatrix} z[N] \\ z[N-1] \\ \vdots \\ z[1] \end{pmatrix} = \left(\frac{1}{\sqrt{N}}W_N\right)^{-1} D_H \left(\frac{1}{\sqrt{N}}W_N\right)\begin{pmatrix} ss[N] \\ ss[n-1] \\ \vdots \\ ss[1] \end{pmatrix} + \begin{pmatrix} w[N] \\ w[N-1] \\ \vdots \\ w[1] \end{pmatrix}$$

$$\tilde{S}\tilde{S} \stackrel{\Delta}{=} \begin{pmatrix} \tilde{s}\tilde{s}[N] \\ \tilde{s}\tilde{s}]n-1] \\ \vdots \\ \tilde{s}\tilde{s}[1] \end{pmatrix} = \left(\frac{1}{\sqrt{N}} W_N \right) \begin{pmatrix} ss[N] \\ ss[n-1] \\ \vdots \\ ss[1] \end{pmatrix} \tag{1.22}$$

$$\tilde{z} \stackrel{\Delta}{=} \begin{pmatrix} \tilde{z}[N] \\ \tilde{z}[n-1] \\ \vdots \\ \tilde{z}[1] \end{pmatrix} = \left(\frac{1}{\sqrt{N}} W_N \right)^{-1} \begin{pmatrix} \tilde{z}[N] \\ \tilde{z}[n-1] \\ \vdots \\ \tilde{z}[1] \end{pmatrix} \tag{1.23}$$

So we get the relationship between the source data and the channel output vector after DFT:

$$\tilde{z} = D_H \tilde{s} s + \tilde{w} \tag{1.24}$$

Because D_H is diagonal, this matrix product is the same as an element-wise multiplication:

$$\tilde{z}[n] = H[n]\tilde{s}s[n] + \tilde{w}[n] \qquad n = 1, \ldots, N \tag{1.25}$$

We now have N parallel subchannels, each of which is a scalar channel with gain H[m]. Each vector of N data symbols in OFDM transmission is known as an OFDM frame of symbol. Each sub-channel H[n] represents a subcarrier.

But this added CP is not entirely free, it should be noticed that it comes with both power and bandwidth penalty since "r" redundant symbols are sent, where the data rate and power losses are

$$\text{Power Loss} = \text{Rate loss} = \frac{L}{L+r} \tag{1.26}$$

To decrease this inefficiency, the number of subcarriers can be increased to achieve L>>r, however other sacrifices must be mode as L gets large, so desirable properties is traded off as with most system design problems.

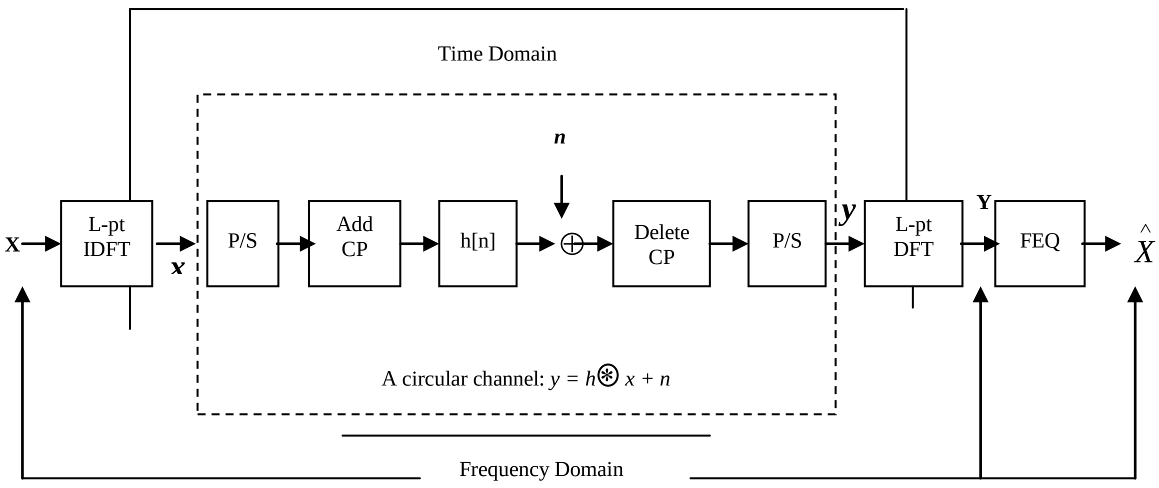


Figure 1.1 OFDM block diagram

<u>Timing and frequency synchronization</u>

In order to demodulate an OFDM signal, two essential synchronization tasks must be performed at the receiver:

1- Timing synchronization, where the timing of Best of the symbol, and the optimal timing instants must be.

2- Frequency Synchronization, where the receiver must align its carrier frequency as closely as possible with the transmitted carrier frequency.

The IFFT effectively modulates each dates symbol onto a unique carrier freq. where the transmitted signal is the superposition of all the individual carriers. For a "since" function form subcarrier in the frog the frequency response of each subcarrier becomes a "since" function. In the frequency domain, with a perfect carrier frequency synchronization, the receiver samples at the peak of each subcarrier, where the desired subcarrier amplitude is maximized, and the intercarrier interference (ICI) is zero.

Timing synchronization

The effect of timing errors in symbol synchronization is some what released in OFDM and that is due to the (cp).

As perilously as summed that only the "L" time domain samples after the CP are used by the receiver gives the result of "Perfects" timing synchronization. Even if the CP length is equivalent to the length of the channel impulse response r, successive OFDM symbols can be decoded ISI free.

In the case of imperfect synchronization, it will still be possible to tolerate a timing offset of r seconds without performance degradation as $o \le t \le Tm - Tg$[2], where Tm is the maximum channel this relation determines the acceptable range of y which is referred to as "Timing Synchronization margin". In this context wherever y<o then sampling is earlier than at ideal time instant, meanwhile if τ >o sampling occurs later than at the ideal instant. On the other hand if the offset is not within this window $o \le t\underline{L}Tm - Tg$, ISI occurs.

Considering τ >o, result in a receiver loss of some of the desired energy due to the delayed version reception of the early samples X_0, X_1, …. in addition to undesired energy from the subsequent symbol. Meanwhile for τ <Tm-Tg, there is also some energy lost but with interference from the preceding symbol this time. For both scenarios the signal to noise ratio (SNR) loss can be approximated as [2].

$$\Delta SNR(\tau) \approx -2\left(\frac{\tau}{LT_s}\right)^2 \tag{1.27}$$

Frequency Synchronization

OFDM achieves a higher degree of bandwidth efficiency than do other wideband systems. The subcarrier packing is extremely tight compared to conventional modulation techniques, which require a guard band on the order of 50 percent or more. The price to be paid for this bandwidth efficiency is that the multicarrier signal is very sensitive to frequency offsets, owing to the fact that the subcarriers overlay rather than

having each subcarrier truly spectrally isolated.

A detailed mathematical analysis of frequency misalignment is produced in chapter two, in the context of a very important causing factor of it, which happens to be the Doppler Effect.

Obtaining Synchronization in OFDM systems

Many synchronization algorithms have been developed in the topic. Generally, the methods can be categorized as based on either pilot symbol or blind—cyclic prefix.

In the first category, known pilot symbols are transmitted. Since the receiver knows what

was transmitted, attaining quick and accurate time and frequency synchronization is easy, but at the cost of surrendering some throughput. For example, in the WiMAX system downlink, the preamble consists of a known OFDM symbol that can be used to attain initial synchronization. In the WiMAX uplink, the periodic ranging can be used to synchronize. Since WiMAX is an OFDM-based system with many competing users, each user implements the synchronization at the mobile. This requires the base station to communicate the frequency offset to the MS.

Blind means that pilot symbols are not available to the receiver, so in the second category, the receiver must do the best it can without explicitly being able to determine the effect of the channel. In the absence of pilot symbols, the cyclic prefix, which contains redundancy, can also be used to attain time and frequency synchronization. This technique is effective when the number of subcarriers is large or when the offsets are estimated over a number of consecutive symbols. The principal benefit of CP-based methods is that pilot symbols are not needed, so the data rate can nominally be increased. In WiMAX, accurate synchronization, and especially channel estimation, are considered important enough to warrant the use of pilot symbols, so the blind

techniques are not usually used for synchronization. They could be used to track the channel in between preambles or ranging signals, but the frequency and timing offsets generally vary slowly enough that this is not required.

(2) MULTI-INPUT MULTI-OUTPUT (MIMO) SYSTEMS

The previous section explained how a system can achieve frequency diversity through the use of multicarrier modulation, and particularly, how OFDM systems achieved that. The focus of this section is spatial diversity, which is created without using the additional bandwidth that time and frequency both require. This is in turn achieved by the usage of antenna arrays or simply multiple antennas both at the transmitter and the receiver, which is referred to as multiple/input multiple/output (MIMO) communication. Despite the cost associated with additional antenna elements and their accompanying radio frequency chains, the gain from antenna arrays is enormous.

Benefits of Spatial Diversity:

As mentioned, one main advantage of spatial diversity is that no additional bandwidth or power is needed to take advantage of this concept in comparison with time and frequency diversity. Although the additional cost cannot be negligible, this trade-off often maintains its worth, since it was found that only one additional antenna appropriately spaced is sufficient to eliminate most deep fades.

<u>1-Array Gain</u>

Two types of gain appears in the presence of multiple antennas at the receiver:

a. Diversity gain, that is the gain resulting from the multiple independent channels created between the transmitter and the receiver.

b. Array gain, that is the gain achieved by coherently combining the energy received by each of the antennas.

It was proven that the received SNR increases linearly with the number of receive antennas N_r, even in the case where the channels are completely correlated as in line-of site systems[2,5,7], the case where no diversity gain exists.

In a $N_t \times N_r$ system, where N_t is the number of transmit antennas and N_r is the number of receive antennas. The array gain is N_r.

For a correlated flat fading channel with 1x N_r system, the received signal at each antenna is written as

$$y_i = h_i x + n_i = hx + n_i \tag{1.28}$$

Where $h_i = h$ due to correlation.

The SNR on a single antenna is

$$\gamma_i = \frac{|h|^2}{\sigma^2} \tag{1.29}$$

Where σ^2 is the noise power assuming signal energy of unity ($\varepsilon_x = E|x|^2 = 1$).

Adding all the receive antenna paths gives

$$y = \sum_{i=1}^{N_r} y_i = N_r hx + \sum_{i=1}^{N_r} n_i \tag{1.30}$$

Hence, the combined SNR for uncorrelated noise per branch is

$$\gamma_\Sigma = \frac{|N_r h|^2}{N_r \sigma^2} = \frac{N_r |h|^2}{\sigma^2} \tag{1.31}$$

Which assures the previous conclusion.

2-Diversity Gain and Decreased Error Rate

In additive noise, the bit error probability (BEP) can be written for any modulation scheme as [2,9]

$$P_b \approx c_1 e^{-c_2\gamma} \tag{1.32}$$

Where c_1 and c_2 are constants that depend on the modulation type, and γ is the received SNR.

In fading, without diversity, the average BEP can be written as

$$\overline{P}_b \approx c_3\gamma^{-1} \tag{1.33}$$

From those two relations it is obvious that the BEP with fading is dramatically worse.

Now if the presence of N_t transmit antennas and N_r receive antennas is considered (where the antennas are spaced efficiently), the "Diversity order" of the system is $N_d = N_t \times N_r$, which represents the number of uncorrelated channel paths between the transmitter and the receiver. Accordingly the average BEP becomes [2]

$$\overline{P}_b \approx c_4\gamma^{-N_d} \tag{1.34}$$

This is an enormous improvement. On the other hand, the case in which the antennas are insufficiently spaced or the channel is LOS, i.e., only an array gain is possible, the average BEP will decrease to [2]

$$\overline{P}_b \approx c_5(N_d\gamma)^{-1} \tag{1.35}$$

This shows that only a linear increase in SNR is provided by array gain.

3- Increased Data Rate

Receive diversity techniques also increase the average received SNR linearly at best. The Shannon capacity formula which gives the maximum achievable data rate of a single communication link in additive white Gaussian noise (AWGN) as

$$C = B\log_2(1+\gamma) \quad (1.36)$$

Where C is the capacity, or maximum error-free data rate, B is the bandwidth of the channel and γ is the SNR (or SNIR). With sufficient diversity and/or sufficient coding schemes it may possible in some wireless channels, to approach the Shannon limit.

It is obvious that the capacity increases logarithmically with respect to the number of antennas, resulting in a less increase of the capacity at higher SNRs. For an enhanced data increase at higher SNRs, the multiantenna channel can instead be used to send multiple independent streams.

Receive Diversity

The most prevalent form of spatial diversity is receive diversity, often with only two antennas. Receive diversity places no particular requirements on the transmitter but requires a receiver that processes the Nr received streams and combines them in some fashion. Two of the widely used combining algorithms are: selection combining (SC) and maximal ratio combining (MRC). Although receive diversity is highly effective in both flat fading and frequency-selective fading channels, we focus on the flat-fading scenario, in which the signal received by each of the antennas is uncorrelated and has the same average power.

<u>a. Selection Combining</u>

It is the simplest type of combiner, as it simply estimates the instantaneous strengths of each of the streams and selects the highest one. Since it ignores the useful energy on the other streams, SC is clearly suboptimal, but its simplicity and reduced hardware requirements make it attractive.

The diversity gain from using selection combining can be confirmed quite quickly by considering the outage probability, defined as the probability that the received SNR drops below some required threshold, . Assuming uncorrelated receptions of the signal,

$$P_{out} = P[\gamma_1 < \gamma_0,\ \gamma_2 < \gamma_0,\ ..., \gamma_{N_r} < \gamma_0]$$

$$= P[\gamma_1 < \gamma_0]P[\gamma_2 < \gamma_0]...\ P[\gamma_{N_r} < \gamma_0]$$

$$= P^{N_r} \qquad (1.37)$$

For a Rayleigh fading channel,

$$p = 1 - e^{-\gamma_0 / \bar{\gamma}} \qquad (1.38)$$

Where $\tilde{\gamma}^{r}$ is the average received SNR at that location—for example, owning to pathloss. Thus, selection combining dramatically decreases the outage probability to

$$P_{out} = (1 - e^{-\gamma_0 / \bar{\gamma}})^{N_r} \qquad (1.39)$$

The average received SNR for N_r branch SC can be derived in Rayleigh fading to be

$$\bar{\gamma}_{sc} = \bar{\gamma} \sum_{i=1}^{N_r} \frac{1}{i} \qquad (1.40)$$

$$= \bar{\gamma}(1 + \frac{1}{2} + \frac{1}{3} + ... + \frac{1}{N_r}) \qquad (1.41)$$

Hence, although each added (uncorrelated) antenna does increase the average SNR, it does so with rapidly diminishing returns.

b. Maximal Ratio Combining

Maximal ratio combining combines the information from all the received branches in order to maximize the ratio of signal-to-noise power, which gives it its name. MRC works by weighting each branch with a complex factor $q_i = |q_i| e^{j\phi_i}$ and then adding up the branches. The received signal on each branch can be written as $x(t)h_i$, assuming that the fading is flat with a complex value of $h_i = |h_i| e^{j\theta_i}$ on the ith branch. The combined signal can then be written as S

$$y(t) = x(t)\sum_{i=1}^{N_r} |q_i||h_i| \exp\{ j(\phi_i + \theta_i\} \tag{1.42}$$

If we let the phase of the combining coefficient $\phi_i = -\theta_i$ for all the branches, the signal-to-noise ratio of $y(t)$ can be written as

$$\gamma_{MRC} = \frac{\varepsilon_x (\sum_{i=1}^{N_r} |q_i||h_i|)^2}{\sigma^2 \sum_{i=1}^{N_r} |q_i|^2} \tag{1.43}$$

Where ε_x is the transmit signal energy. Maximizing this expression by taking the derivative with respect to $|q_i|$ gives the maximizing combining values as $|q_i^*|^2 = |h_i|^2 / \sigma^2$; that is, each branch is multiplied by its SNR. In other words, branches with better signal energy should be enhanced, whereas branches with lower SNRs should be given relatively less weight. The resulting signal-to- noise ratio can be found to be

$$\gamma_{MRC} = \frac{\varepsilon_x \sum_{i=1}^{N_r} |h_i|^2}{\sigma^2} = \sum_{i=1}^{N_r} \gamma_i \tag{1.44}$$

MRC is intuitively appealing: The total SNR is achieved by simply adding up the branch SNRs when the appropriate weighting coefficients are used. It is noted that although MRC does in fact maximize SNR and generally performs well, it may not be optimal in many cases since it ignores interference powers the statistics of which may differ from branch to branch.

Equal gain combining (EGC), which corrects only the phase and hence as the name of the

technique suggests uses $|q_i|$=1 and $\phi_i = -\theta_i$ for all the combiner branches, achieves a postcombining

SNR of

$$\gamma_{EGC} = \frac{\varepsilon_x \sum_{i=1}^{N_r} |h_i|^2}{N_r \sigma^2} \tag{1.45}$$

The most notable difference between the last two equations is that EGC incurs a noise penalty in trade for not requiring channel gain estimation. EGC is hence suboptimal compared to MRC, assuming that the MRC combiner has accurate knowledge of , particularly when the noise variance is high and there are several receive branches. For an interference-limited cellular system, such as WiMAX, MRC would be strongly preferred to either EGC or SC.

(3) OFDMA

OFDMA stands for "orthogonal frequency division multiple access", the system which occupies the title of this search. OFDMA is simply a multi-user OFDM system, hence all previous discussions held about OFDM stand for OFDMA as well, the fact that this system is a multi-user system, i.e. multiple inputs goes through or access the system and

multiple outputs come out of it, implies that OFDMA is basically a MIMO system as well too.

It is a flexible multiple-access technique that can accommodate many users with widely varying applications, data rates and QoS requirements. It can also be visualized as a hybrid of FDMA and TDMA where users are dynamically assigned subcarriers (FDMA) in different time slots (TDMA). This multiple access is performed before the IFFT operation i.e. in the digital domain, hence dynamic and efficient bandwidth allocation is possible, allowing the integration of sophisticated frequency and time scheduling algorithms for the best service of users.

Recalling the MIMO system equations given in the previous section which are applied here, the MIMO-OFDMA channel matrix (for a subcarrier n and user k) in the mentioned equations becomes as follows:

$$\overline{\overline{H}}_{k,n} = \begin{pmatrix} h_{1,1} & h_{1,2} & \cdots & h_{1,N_R} \\ h_{2,1} & & & . \\ & & & . \\ h_{N_T,1} & & \cdots & h_{N_T,N_R} \end{pmatrix} \tag{1.46}$$

So the Overall channel matrix becomes:

$$\overline{\overline{H}} = \begin{pmatrix} \overline{\overline{H}}_{1,1} & \overline{\overline{H}}_{1,2} & & \cdots & \overline{\overline{H}}_{1,N} \\ \overline{\overline{H}}_{2,1} & & & & \overline{\overline{H}}_{2,N} \\ . & . & & & . \\ . & & . & \overline{\overline{H}}_{k,n} & . \\ . & & & & . \\ \overline{\overline{H}}_{K,1} & & & \cdots & \overline{\overline{H}}_{K,N} \end{pmatrix} \tag{1.47}$$

The eigenvalue of $\overline{\overline{H}}^{H}_{k,n}.\overline{\overline{H}}^{H}_{k,n}$ *is* $\{\lambda^{(i)}_{k,n}\}_{i=1,\ldots,M_{k,n}}$. And the number of channel elements are

$K \times N \times N_T \times N_R$

Multi-user Diversity and Adaptive Modulation

These are the two key principle that enable high performance in OFDMA system.

Multi-user diversity describes the gains available by selecting a user or a subset of user hawing “good conditions”.

a- Multi-user diversity:

Adaptive subcarrier allocation in OFDMA is motivated to exploit multi-user diversity. Since OFDMA systems have a number of subcarriers, here the focus here is on the allocation of a single subcarrier among multiple users.

For a K-user system and a single subcarrier experiencing i.i.d Rayleigh fading (i.e. users' channel gains $(h_k)_s$, are independent of each other), the probability density function (PDF) of user k’s channel gain $p(h_k)$ is given by:

$$P(h_k) = \begin{cases} 2h_k e^{-h_k^2} & if\ h_k \geq 0 \\ 0 & if\ h_k < 0 \end{cases} \tag{1.48}$$

Now if the base station transmits only to the user with the highest channel gain (h_{max}), where $h_{max} = \max\{h_1, h_2, \ldots, h_K\}$ then,

$$P(h_{max}) = 2kh_{max}(1 - e^{-h_{max}^2})^{k-1} e^{-h_{max}^2} \tag{1.49}$$

b-Adaptive Modulation and Coding

In order to take advantage of random fluctuations in wireless communication channels, OFDMA systems use adaptive modulation and coding (AMC).

AMC changes the (coding scheme and/ or the (modulation method) according to the channel – state information, choosing it in a way that it squeezes the most out of the channel’s transmission ability.

The idea behind that is to achieve the highest spectral efficiency at all times by dynamically adapting the modulation and coding schemes to the channel conditions [13].

This is applied simply by transmitting at the highest possible data rate when the channel is good, and transmitting at lower rates when the channel is poor, which contributes to the avoidance of data packet loss.

And that in turn is achieved by using larger constellation such as 64 QAM and higher rate (less robust) error-correcting codes, like turbo codes, rate 3/4 convolutional codes…etc for high data rates.

Fifty two configurations of modulation order and coding types and rates are possible, those constellation/ coding sets are called “Burst profiles”.

In most implementations of OFDMA only a fraction of these burst profiles are offered, for example by using six of the common WiMAX burst profiles a large range of spectral efficiencies could be achieved [3].

The above discussion implies that the “feedback channel” is critical for AMC, since the transmitter needs to know the “channel SINR" or "γ", where SINR stands for the channel signal to interference plus noise ratio and is given by:

$$\gamma \triangleq \frac{\gamma_r}{P_t} \tag{1.50}$$

Where γ_r is the received SINR, and P_t is the transmit power.

Following Shannon’s Formula:

$$C = \log_2 (1 + SNR) \tag{1.51}$$

Which implies that the throughput increases with increasing SINR, the achieved throughput (capacity) normalized by the BW is defined as [3].

$$T \triangleq (1 - BLER) r \log_2 (M) \quad bps/Hz \tag{1.52}$$

Where r is the coding rate with $r \le 1$, M is the number of points in the constellation, and BLER is the block error rate, which in turn should be monitored as type final word, which in turn should be on whether the data rate should be increased (when BLER is low) or decreased to a more robust setting.

So as a result, the key challenge in AMC is to efficiently control three quantities at once: the transmit power, the transmit rate (constellation), and the coding rate.

Resource – Allocation Techniques for OFDMA

In order to take advantage of multiuser diversity and adaptive modulation discussed earlier, a number of algorithms where created. Those algorithms are used for determining which user to schedule, how to allocate subcarriers among those users, and how to determining the appropriate power levels for each user on each subcarrier.

Users estimate and feedback the channel state information (CSI) to a centralized base station where subcarrier and power are determined according to users's CSI and resource-allocation procedure, once the subcarriers are determined, each user must be informed by the base station with the subcarrier that has been allocated to it. Further this subcarrier mapping must be broadcast to all users whenever the resource allocation changes.

The resource allocation is "constrained optimization problem" for it either minimizes the transmitted power with a constraint on the user data rate or maximizes the total data rate with a constraint on the p_t, which is more relevant to OFDMA systems and is under focus here, assuming that the (BS) has obtained perfect instantaneous (CSI) for all users.

1- Maximum Sum Rate Algorithm

The objective of the maximum sum rate (MSR) algorithm is to maximize the sum rate of all users given a total transmit power constraint [3].

This algorithm is best when the goal is to get as much data as possible through the system but that comes with a drawback, that is, few users which are close to the base station and having excellent channels will be allocated all the system resources.

If $P_{k,l}$ represents user k's transmit power in subcarrier *l*. then the signal-to – interference-plus-noise-ratio for user k in subcarrier *l*, is expressed as [3].

$$SINR_{k,l} = \frac{P_{k,l} h_{k,l}^2}{\sum_{j=1, j\neq k}^{K} p_{j,l} h_{k,l}^2 + \sigma^2 \frac{B}{L}} \tag{1.53}$$

K Number of users.

L: Number of subcarriers.

$H_{k,l}$: Envelope of channel gain for user k in subcarrier l

$P_{k,L}$: Transmit power allocated for user k in subcarrier l

σ^2 : AWGN power spectrum density.

P_{tot} : Total transmit power available at the base station.

B : Total transmission bandwidth.

According to shannon capacity formula, the MSR algorithm maximizes the following quantity:

$$\max \sum_{k=1}^{K} \sum_{l=1}^{L} \frac{B}{L} \log(1 + SINR_{k,l}). \tag{1.54}$$

With the total power constraint $\sum_{k=1}^{K} \sum_{l=1}^{L} P_{k,l} \leq P_{tot}$.

Since the sum capacity is maximized if the total throughput in each subcarrier is maximized, then this maximum sum problem can be splitted into L simpler problems, one for each subcarrier. Then the sum capacity in subcarrier *l*, is given by [3].

$$C_{\ell} = \sum_{k=1}^{K} \log\left(1 + \frac{P_{k,l}}{P_{tot,l} - P_{k,l} + \frac{\sigma^2}{h_{k,l}^2}\frac{B}{L}}\right) \tag{1.55}$$

Where $P_{tot,l} - P_{k,l}$ is the other users interference to user k in subcarrier l.

It is obvious that C_l is maximum when all power on l is assigned to user a single user k which has the best gain on that channel, and hence this algorithm is sometimes called "greedy optimization".

2-Maximum Fairness Algorithm

The target of this algorithm is to allocate the subcarriers and power such that the minimum user's data rate is maximized, or in other words to equalize the data rates of all users, and hence sometimes referred to as a "max-min" problem.

The need of such an algorithm arises from the fact that in cellular systems such as WiMAX-OFDMA pathloss attenuation varies in several orders of magnitude between users, making some users extremely underserved by the previously discussed MSR procedure.

But this allocation procedure is much more difficult than MSR, since the simultaneous allocation of the optimum power and subcarriers is considerably difficult. Therefore, the need arises for low-complexity, suboptimal algorithms, in which separate subcarrier and power allocation are performed.

For example, an approach to this idea is to assume initially "equal power" allocation to each subcarrier, then to perform alterative subcarrier allocation to low-rate user with the

best channel on it, once this task is completed an optimum water- filling power allocation can be performed, that is simply to allocate more power to strong channels and less power to weak channels, a method which will be discussed in more detail in the following chapter.

3-Proportional Rate Constraints Algorithm

The previous method is a suboptimal method, that is due to the fact that the total throughput is limited largely by the worst SINR user, since most of the resources are allocated to him.

In a wireless broadband network, different users require application-specific data rates which vary substantially. The proportional rate constraints (PRC) algorithm came to serve this requirement as a generalization of the previous algorithm. That is achieved by maximizing the sum throughput with the additional constraint that each user's data rate is proportional to a set of predetermined parameters $\{\beta_k\}_{k=1}^{K}$.

The proportional data rates are [3].

$$\frac{R_1}{\beta_1} = \frac{R_2}{\beta_2} = ... = \frac{R_K}{\beta_K}. \qquad (1.56)$$

And each user's achieved data rate R_k is [3].

$$R_k = \sum_{l=1}^{L} \frac{\rho_{k,n} B}{L} \log_2\left(1 + \frac{P_{k,l} h_{k,l}^2}{\sigma^2 \frac{B}{L}}\right) \qquad (1.57)$$

$\rho_{k,l}$ can be of the value of either 1 or o, since subcarrier l is either used or not by user k. If $\beta_k = 1 \, \forall \, k$, then it would be the same setup as the maximum fairness algorithm, meanwhile for varying β_k values, any arbitrary data rates can be achieved.

This PRC optimization problem is very difficult to solve in general, since it involves both binary variables $\rho_{k,l}$ and continuous variables $P_{k,l}$.

4-Proportional Fairness Scheduling

This algorithm adds flexibility to the three previous scheduling algorithms -which attempt to “instantaneously” achieve each of their objectives-by allowing a third element to enter the trade-off and that is: latency, in order to achieve the previous three objectives over “time”.

Large latencies, i.e. in order of seconds are not acceptable, therefore the need for scheduling algorithms that balances through put and latency stems and achieves some degree of fairness.

If R_k (t) denotes the instantaneous data rate that user R can achieve at time t, and T_k (t) is the average throughput for user k up to time slot t, then the PF scheduler selects the user, K^* with the highest $R_k(t)/T_k(t)$ for transmission.

The average throughput $T_k(t)$ for all users is then up dated according to [3].

$$T_k(t+1) = \begin{cases} (1-\frac{1}{t_c})T_k + \frac{1}{t_c}R_k(t) & K = K^* \\ (1-\frac{1}{t_c})T_k(t) & K \neq K^* \end{cases} \tag{1.58}$$

Since the selection goes toward the user with the largest instantaneous data relative to its average throughput, then “bad” channels for each user aren’t likely to be selected.

The parameter controlling the latency of the system is t_c. The latency of the system increases if it is large, with the benefit of higher sum throughput. The latency decreases if t_c is small.

Since the values of the average throughput change more quickly, at some throughput expense.

This algorithm was originally designed for a single-channel time-slotted system, it can also be adapted to an OFDMA system. Although this system multiple users can transmit on different subcarriers simultaneously, the PF algorithm is extended by treating each sub carrier independently. So if $R_k\ (t, n)$ denotes the supportable data rate for user *k* in subcarrier *n* at time slot *t*.

Then for each subcarrier the user with the largest R_k (t, n)/ T_k (t) is selected for transmission, then the average user through put is updated as [3].

$$T_k(t+1) = (1 - \frac{1}{t_c})T_k(t) + \frac{1}{t_c} \sum_{n \in \Omega_k(t)} R_k(t,n) \qquad (1.59)$$

For k = 1, 2, ……., K. where Ω_k (t) is the set of subcarriers user k is scheduled for transmission at time slot t.

All the end of this section, it is worth mentioning that scheduling algorithms do not need to be specified by OFDMA systems - and WiMAX, is one of them - and they are really not. On the other hand several key contributes must be standardized, and those include subchannelization, mapping messages, and ranging which are out of the scope of this research.

Actually WiMAX users are allocated blocks of subcarriers rather than individual subcarriers in order to lower the complexity of the subcarrier-allocation-algorithm and the mapping messages, which are broadcast by the BS to inform each MS with the subcarriers that are intended for it.

CHAPTER TWO

EFFECTS OF MOBILITY ON OFDMA SYSTEMS

The goal of this chapter is to explain the fundamental factors affecting the received signal in a wireless system. Large-scale fading and small-scale fading are the main factors causing received power to vary over distances in wireless channels. Consequently, those affect the performance measurements of the system, mentioning in particular the system's "Capacity" in more detail, since occupying the title of this thesis.

Achieving high data rates in terrestrial wireless communication is difficult. High data rates for wireless local area networks, namely the IEEE 802.11 family of standards, became commercially successful only around 2000. Despite many promising technologies, the reality of a wide area network that services many users at high data rates with reasonable bandwidth and power consumption, while maintaining high coverage and quality of service, has not yet been achieved.

1- Large-Scale Fading.

a. Pathloss

Pathloss (or path attenuation) is the reduction in power density (attenuation) of an electromagnetic wave as it propagates through space. It is a major component in the analysis and design of the link budget of a communication system.

As pathloss accounts for the distance-dependent relationship between transmitted and received power, it represents signal level attenuation caused by free space propagation, reflection, diffraction and scattering. [5]

Path loss is also influenced by terrain contours, environment (urban or rural, vegetation and foliage), propagation medium (dry or moist air), the distance between the transmitter and the receiver, and the height and location of antennas.

The free-space pathloss formula, or Friis formula, is given more precisely as [3,5],

$$P_r = P_t \frac{\lambda^2 G_t G_r}{(4\pi d)^2} \tag{2.1}$$

- P_r and P_t are the received and transmitted powers respectively.
- G_r and G_t are the gains of the transmitting and receiving antennae respectively.
- λ is the wavelength.
- d is the distance between the transmitting and receiving antennae in meters.

Pathloss is generally the most important parameter predicted by large-scale propagation models, many models where developed for different types of path loss, three kinds of models exist, empirical models, semi-deterministic models, and deterministic models

In order to more accurately describe various propagation environments, empirical models are often developed using experimental data.

One of the simplest and most common is the empirical path loss formula[5]:

$$P_r = P_t P_o \left(\frac{d_o}{d}\right)^{\alpha} \tag{2.2}$$

This formula groups all the various effects into two parameters: the pathloss exponent α and the measured pathloss P_o at a reference distance of d_o, which is often chosen as 1 meter. P_o should be determined from measurements, and it is often well approximated, within several dB, for example $P_o = (4\pi/\lambda)^2$ when $d_o = 1$. This simple empirical pathloss formula is capable of reasonably representing most of the important pathloss trends with only these two parameters.

b. Shadowing

Many factors other than distance can have a large effect on the total received power. For example, obstacles located between the transmitter and the receiver such as trees and buildings may cause temporary degradation in received signal strength.

Since modeling the locations of all objects in different communication environments is generally impossible, the standard method of accounting for these variations in signal strength is to introduce a random effect called shadowing. With shadowing, the empirical pathloss formula becomes [3]:

$$P_r = P_t P_o \chi \left(\frac{d_o}{d} \right)^{\alpha} \tag{2.3}$$

Where χ is a sample of the shadowing random process. Hence, the received power is now also modeled as a random process. In effect, the distance trend in the pathloss can be thought of as the mean, or expected, received power, whereas the shadowing value causes a perturbation from that expected value. Since shadowing is caused by macroscopic objects, it typically has a correlation distance on the order of meters or tens of meters. Hence, shadowing is often alternatively called large-scale fading.

The shadowing value is typically modeled as a lognormal random variable, that is,

$$\chi = 10^{\frac{x}{10}}, where \;\; x \approx N(0, \sigma_s^2), \tag{2.4}$$

Where $N(0, \sigma_s^2)$ is a Gaussian (normal) distribution with mean 0 and variance σ_s^2. With this formulation, the standard deviation is expressed in dB. Typical values for are in the 6–12 dB range.

Shadowing has an important effect in wireless networks because it causes the received signal- to- interference +noise ratio (SINR) to vary dramatically over long time scales. In some locations in a given cell, reliable high-rate communication may be nearly

impossible. Although shadowing can sometimes be beneficial—for example, if an object is blocking interference—it is generally detrimental to system performance.

2- Small-Scale Fading.

Meanwhile pathloss and shadowing show up due to distance and obstacles, “small-scale fading”, or simply “fading” is caused by the reception of multiple versions of the same signal. Those multiple received versions are caused by reflections which are known as multipath.

Depending on the phase difference between the arriving signals, the interference can be either constructive or destructive, which causes a very large difference in the amplitude of the received signal even over very short distances.

Doppler Effect

Besides the multipath propagation, the Doppler Effect has a negative influence on the transmission characteristics of the mobile radio channel. Due to the movement of the mobile unit, the Doppler Effect causes a frequency shift of each of the partial waves.

The angle of arrival (AoA), is defined by the direction of arrival of the nth incident wave and the direction of motion of the mobile unit. It determines the Doppler frequency (frequency shift) of the nth incident wave according to the relation:

$$f_D = f_{DMAX} \cos\theta \qquad (2.5)$$

Where θ is the angle of arrival, f_D is the Doppler frequency shift, and f_{DMAX} is the maximum Doppler frequency shift defined as [7]:

$$f_{DMAX} = \frac{v}{c} f_0 \qquad (2.6)$$

Where v is the speed of motion of the mobile unit, f_0 is the carrier frequency, and c is the speed of light which equals $3x10^8 m/s$.

It is obvious that if transmitter and the receiver are moving fast relative to each other and hence the Doppler is large, the channel will change much more quickly than if the transmitter and the receiver are stationary.

Additionally, the large Doppler at high mobility and frequency can also degrade the OFDM subcarrier orthogonality, as mentioned in Chapter 1.

The following analysis is stated to show the effect of a general frequency shift on subcarriers.

Consider subcarriers of the "sinc" form. The sinc function is defined as

$$\sin c(x) = \frac{\sin(\pi x)}{\pi x} \tag{2.7}$$

With this definition, it can be confirmed that sinc(0) = 1and that zero crossings occur at ± 1, ±2, ±3, .. Sinc functions occur commonly because they are the frequency response of a rectangular function. Since the sine waves existing in each OFDM symbol are truncated every T seconds, the width of the main lobe of the subcarrier sinc functions is 2/T, i.e., there are zero crossings every 1/T Hz.

Practically, the frequency offset is not always zero. The major causes for this are mismatched oscillators at the transmitter and the receiver and Doppler frequency shifts owing to mobility. This will degrade the orthogonality of the received sign since the received samples of the FFT will contain interference from the adjacent subcarriers. We now analyze this intercarrier interference in order to better understand its effect on OFDM performance which is our case of study in this thesis.

The matched filter receiver corresponding to subcarrier l can be simply expressed for

the case of rectangular windows, neglecting the carrier frequency, as [2]

$$x_l(t) = X_l e^{j\frac{2\pi l t}{LT_s}} \tag{2.8}$$

Where *1/LTs* = Δf, and again *LTs* is the duration of the data portion of the OFDM symbol: *T* = *Tg* + *LTs*. An interfering subcarrier *m* can be written as

$$x_{l+m}(t) = X_m e^{j\frac{2\pi(l+m)t}{LT_s}} \tag{2.9}$$

If signal is demodulated with a fractional frequency offset of δ, $|\delta| \le \frac{1}{2}$.

$$\hat{x}_{l+m}(t) = X_m e^{j\frac{2\pi(l+m+\delta)t}{LT_s}} \tag{2.10}$$

The ICI between subcarriers *l* and *l* + *m* using a matched filter, the FFT, is simply the inner product between them [2]

$$I_m = \int_0^{LT_s} x_l(t)\hat{x}_{l+m}(t)dt = \frac{LT_s X_m (1 - e^{-j2\pi(\delta+m)})}{j2\pi(m+\delta)} \tag{2.11}$$

It can be seen that in this equation, if $\delta=0$ then *Im=0*, if and *Im=0*, then $\delta=0$. The total average ICI energy per symbol on subcarrier *l* is then

$$ICI_l = E[\sum_{m\neq l}|I_m|^2] \approx C_0(LT_s\delta)^2\varepsilon_x \tag{2.12}$$

Where *Co* is a constant that depends on various assumptions and ε_x is the average symbol energy. The approximation sign is used because this expression assumes that there are an infinite number of interfering subcarriers. Since the interference falls off quickly with *m*, this assumption is very accurate for subcarriers near the middle of the band and is pessimistic by a factor of 2 at either end of the band.

Now going back to Doppler Effect, the Doppler frequency shift could be substituted in the last equation, in a way to realize the effect on subcarriers.

The SNR loss induced by frequency offset is given by [2]

$$\Delta SNR = \frac{\varepsilon_x / N_0}{\varepsilon_x /(N_0 + C_0(LT_s\delta)^2} \quad (2.13)$$

CHAPTER THREE

CAPACITY AND CAPACITY ENHANCEMENT

Capacity

The discussed phenomena in the previous chapter will be reflected on the wireless communication system's performance. Since the "Transmission Capacity" or simply "Capacity", is one of the most important system performance measures and indicators, it occupied the title of this thesis. Hence a detailed discussion of this term is stated here throughout the following pages.

The "Transmission Capacity" is physically defined as the maximum theoretical transmission rate of data such that the probability of error is arbitrarily small.

MIMO System Model

Consider a MIMO system with a transmit array of M_T antennas and a receive array of M_R antennas. The transmitted matrix is a $M_T x1$ column matrix ***s*** where s_i is the ith component transmitted from antenna *i*. If the channel is considered to be a Gaussian channel, the elements of ***s*** are considered to be independent identically distributed (i.i.d) Gaussian variables. In the case that the channel is unknown at the transmitter, signals transmitted from each antenna are assumed to have equal powers of E_s/M_T. The covariance matrix for this transmitted signal is given by [5].

$$R_{SS} = \frac{E_S}{M_T} I_{M_T} \tag{3.1}$$

Where I_{M_T} is an $M_T x M_T$ identity matrix, and Es is the total transmitted symbol energy irrespective of antennas M_T. Narrowband signal transmission is considered, and flat frequency response is present. The channel matrix **H** is a complex matrix with

dimension $M_T x M_T$, where each component $h_{i,j}$ of the matrix represents the fading coefficient from the *jth* transmit antenna to the *ith* receive antenna. Those elements of **H** can be either deterministic or random.

We assume that each of the receive antennas receives energy equal to the total transmitted energy *Es*, this implies that antenna gains and signal attenuation are ignored. The channel matrix can be estimated at the receiver by transmitting a training sequence. If we require the transmitter know this channel, then this information must be sent to it by a feedback channel.

The noise at the receiver is a column matrix **n** of $M_R x1$ size. The components of this vector are zero mean circularly symmetrical complex Gaussian (ZMCSCG) variables. The covariance matrix of the received noise is given by [5]

$$R_{nn} = E[nn^H] \tag{3.2}$$

In case of uncorrelated components of *n* the covariance matrix becomes.

$$R_{nn} = N_0 I_{M_R} \tag{3.3}$$

Where each transmission branch has identical noise power N_0.

Due to our previous assumption, that the total received power per antenna is equal to the total transmitted power, then the SNR is

$$\gamma = \frac{E_s}{N_0} \tag{3.4}$$

And the received vector, which is a $M_R\ x1$

$$r = Hs + n \tag{3.5}$$

The covariance matrix of the received signal defined as $E[rr^H]$, is given by [5] $R_{rr} = HR_{ss}H^H$.

MIMO System Capacity

The following analysis concerning the MIMO system capacity is taken from references [3,5,9].

Assume that the channel knowledge is known only at the receiver and unavailable at the transmitter.

The capacity of MIMO channel is defined as [9]

$$C = \max_{f(s)} I\ (s;y) \tag{3.6}$$

Where f(s) is the probability distribution of the vector s and I(s;y) is the mutual information between vectors **s** and **y**. we note that [9]

$$\text{I (s; y) = H(y) – H(y}|s) \tag{3.7}$$

Where H(y) is the differential entropy of the vector **y**, while H(y$|s$) is the conditional differential entropy of the vector **y**, given knowledge of the vector **s**. Since the vectors **s** and **n** are independent, H(y$|s$) = H(n). From (3.7),

$$\text{I (s; y) = H (y) – H (n)} \tag{3.8}$$

Maximizing the mutual information I(s;y) reduces to maximizing H(y). The covariance matrix of **y**, $R_{yy} = E[yy^H]$, satisfies [5]

$$R_{yy} = \frac{E_s}{M_T} HR_{ss}H^H + N_0 I_{M_R} \tag{3.9}$$

Where $R_{ss} = E[ss^H]$ is the covariance matrix of **s**. Among all vectors with a given covariance matrix R_{yy}, the differential entropy H(y) is maximized when **y** is maximized when **y** is ZMCSCG. This implies that **s** must also be ZMCSCG vector, the distribution of which is completely characterized by R_{ss}. The differential entropies of the vectors **y** and **n** are given by[5]

$$H(y) = \log_2(\det(\pi e R_{yy}))bps/H_z$$

$$H(n) = \log_2(\det(\pi e \sigma^2 I_{M_R}))bps/Hz \quad (3\text{-}10)$$

Therefore, I (s; y) in (3.8) reduces to

$$I(s;\, y) = \log_2 \det\left(I_{M_R} + \frac{E_s}{M_T N_0} H R_{ss} H^H \right) bps/Hz \quad (3\text{-}11)$$

And from (3.6), the capacity of the MIMO channel is given by [5]

$$C = \max_{Tr(R_{ss})=M_T} \log_2 \det\left(I_{M_R} + \frac{E_s}{M_T N_0} H R_{ss} H^H \right) bps/Hz \quad (3\text{-}12)$$

The capacity *C* in (3.12) is also called error-free spectral efficiency or data rate per unit bandwidth that can be sustained reliably over the MIMO link. Thus if our bandwidth is *W Hz*, the maximum achievable data rate over this bandwidth using MIMO techniques is *WC* bit/s.

Channel Unknown to the Transmitter:

If the channel is unknown to the transmitter, then the vector **s** is statistically independent (i.e., $R_{ss} = I_{M_T}$). This implies that the signals are independent and the power is equally divided among the transmit antennas. The capacity in such a case is [5], from (3-12)

$$C = \log_2 \det\left(I_{M_R} + \frac{E_s}{M_T N_0} H H^H \right) \quad (3\text{-}13)$$

We shall refer to (3.13) as capacity. Now HH^H is an M_R x M_R positive semidefinite Hermitian matrix. The eigenvalue decomposition of such a matrix is given by $Q \Lambda Q^H$, where Q is an M_R x M_R matrix satisfying $Q^H Q = QQ^H = I_{MR}$ and $\Lambda = diag\{\lambda_1 \lambda_2 ... \lambda_{MR}\}$ with $\lambda_i \geq 0$. We assume that the eigenvalues are ordered so that $\lambda_i \geq \lambda_i + 1$, then [5]

$$\lambda_i = \begin{cases} \sigma_i^2, & if\ i = 1,\ 2,\ ...r \\ 0 & if\ i = r,\ r+1,\ ...M_R \end{cases} \quad (3\text{-}14)$$

Where σ_i are the singular values obtained as $\sum = diag\{\sigma_1 \ \sigma_2 ... \sigma_r\}$ from the singular value decomposition of H = U$\sum$ V^H. Then the capacity of the MIMO channel is given by

$$C = \log_2 \det\left(I_{M_R} + \frac{E_s}{M_T N_0} Q \Lambda Q^H \right) \qquad (3\text{-}15)$$

Using det($\mathbf{I}_m$+**AB**) = det ($\mathbf{I}_n$=**BA**) for matrices **A**(m x n) and **B**(n x m) and $Q^H Q = I_{M_R}$, (3.15) simplifies to

$$C = \log_2 \det\left(I_{M_R} + \frac{E_s}{M_T N_0} \Lambda \right) \qquad (3\text{-}16)$$

$$C = \sum_{i=1}^{r} \log_2 \left(1 + \frac{E_s}{M_T N_0} \lambda_i \right) \qquad (3\text{-}17)$$

where *r* is the rank of the channel and λ_i (i=1, 2, ... , r) are the positive eigenvalues of HH^H . Equation (3.17) expresses the capacity of the MIMO channel as a sum of the capacities of r SISO channels, each having a energy gain of λ_i(i=1, 2, ... , r) and transmit energy E_s/M_T.

This means that the technique of multiple antennas at the transmitter and receiver opens up multiple scalar spatial data pipes between the transmitter and receiver. Furthermore, equal transmit energy is allocated to each spatial data pipe. This is (or the case when the channel is unknown at the transmitter.

The squared Frobenius norm of H is defined as $\|H\|_F^2 = Tr(HH^H) = \sum_{i=1}^{M_R} \sum_{j=1}^{M_T} |h_{i,j}|^2$. Frobenius norm is interpreted as the total power gain of the channel.

Also $\|H\|_F^2 = \sum_{i=1}^{M_R} \lambda_i$ where $\lambda_i (i = 1,2,....M_R)$ are the eigenvalues of $\mathbf{HH}^H$. This total power is fixed so that $\|H\|_F^2 = \beta$. Then if the channel matrix is of full rank such that M_T =M_R=M, the capacity *C* in (3.17) is maximized when

$\lambda_i = \lambda_j = \beta / M \ (i, \ \ j = 1, \ 2..., M)$ (keeping in mind, the channel is unknown, so equal energy distribution). To achieve this, $HH^H = H^H H = (\beta / M) I_M$, (i.e., the channel matrix H should be orthogonal). This gives

$$C = M \log_2\left(1 + \frac{\beta E_s}{N_0 M^2}\right) \qquad (3\text{-}18)$$

If the elements of H have ones along the diagonal, then $\|H\|_F^2 = M^2$ and

$$C = M \ \log_2\left(1 + \frac{E_s}{N_0}\right) \qquad (3\text{-}19)$$

The capacity of an orthogonal MIMO channel is therefore M times the scalar channel capacity.

Random Channels

Previous discussions concerned MIMO capacity when the channel is a deterministic channel. Now consider the case when H is chosen randomly according to a Rayleigh distribution in a quasi-static channel. This is a real-life situation, for example, in wireless LANs with high data rates and low fade rates. We assume that the receiver has perfect knowledge of the channel and the transmitter has no knowledge of the channel. Since the channel is random, the information rate associated with it is also random.

CHAPTER FOUR

SIMULATION AND RESULTS

The goal of this chapter is to simulate an OFDMA system performance which is subject to “Doppler effect”, i.e. the systems subcarrier frequencies are shifted due to the mobility of the users.

Assume an OFDMA system containing fifty percent mobile units, and fifty percent fixed units. Assume random location of active users, with a uniform distribution function for the direction of motion. The Doppler angle [0,π], since θ corresponds to –θ. A total of thirty two subcarriers where considered, with 16-QAM modulation scheme.

The simulations for this scenario where performed using Matlab. Three different carrier frequencies where used in this simulation, 2.5GHz, 3.75GHz, and 5.8GHz which are typical values for an OFDMA system. Those frequencies where used at three different speeds, 60km/h, 90km/h, and 120km/h, which represent typical speeds for moving vehicles. At these speeds the effect of Doppler is obvious and better noticed than at lower speeds. Hence, the system’s performance was under observation nine times for nine runs. Each run gives two curves for one specific speed at one specific frequency, one curve by using all the subcarriers in the system, and the other curve by disabling the five worst subcarriers.

At each case, the symbol error rate (SER) was shown using the average SER for all the subcarriers, and then it was shown by disabling the five worst subcarriers, in addition to the ideal channel performance which indicates the system performance in absence of “Doppler effect”.

The five disabled subcarriers are taken out a total of thirty two subcarriers used in this system. That means, in order to enhance the system capacity performance, approximately sixth of the served users are handed-off to a neighboring base station to avoid low QoS afforded by bad subcarriers.

The following figures show the obtained results:

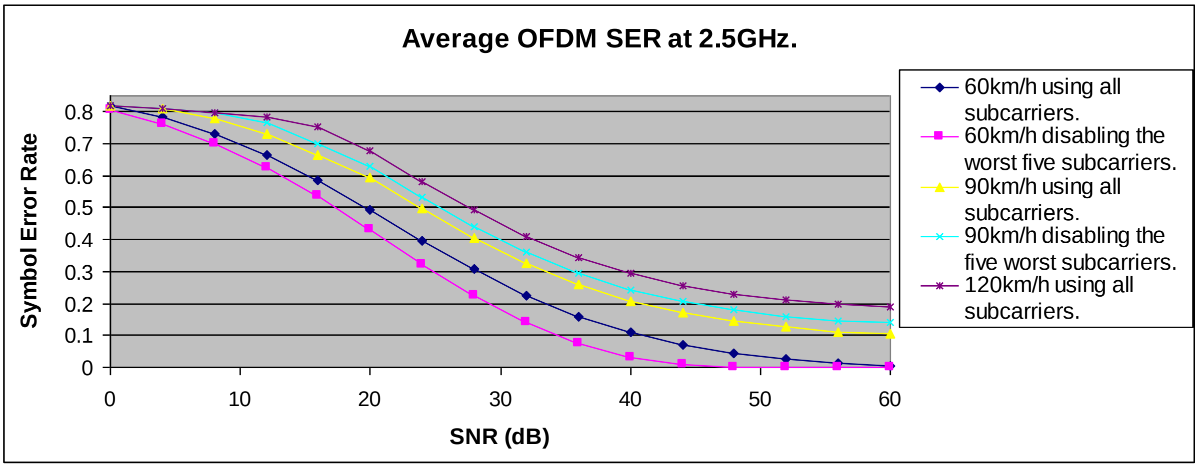


Figure (4-1) System performance at 2.5GHz

Figure (4-1) shows the OFDM system performance at 2.5GHz. This was performed by plotting the SER versus the SNR for the three chosen speeds of motion, 60km/h, 90km/h, and 120km/h respectively. For each speed two plots where obtained as previously mentioned.

From the previous plots the following notes can be extracted:

- As the frequency increases the performance becomes worse, i.e. higher symbol error SER or bit error rate BER for a given SNR. That is trivial according to the previously shown relation:

$$f_{DMAX} = \frac{v}{c} f_0$$

- As the moving speed increases the performance degrades, and that is expected, again according to the Doppler relation, where the maximum Doppler shift increases with the mobile unit's velocity. This is shown in the OFDM SER curves, in the figures corresponding to a given subcarrier frequency.
- A great enhancement in the system's performance at all speeds and all frequencies, when the five worst subcarriers where disabled from the system. That implies a higher quality of service QoS with the existence of Doppler, but of course with a remaining drift from the ideal channel characteristics. To clarify this we can take an example: In figure (4-1) at speed of 60km/h, and for an accepted SER of 0.1, a SNR of 40dB is required when using all subcarriers at the presence of Doppler, meanwhile a SNR value of approximately 36.0dB is needed when the worst subcarriers where avoided. That means a difference of 4dB between the two cases which is a good improvement. This difference also implies less required power values for the same BER.
- At speed of 90km/h and for an accepted SER value of 0.2, a SNR of 45.5dB is needed in the presence of all subcarriers. Meanwhile only 40dB is needed when the bad conditioned subcarriers were avoided, i.e. an improvement of 5.5dB.

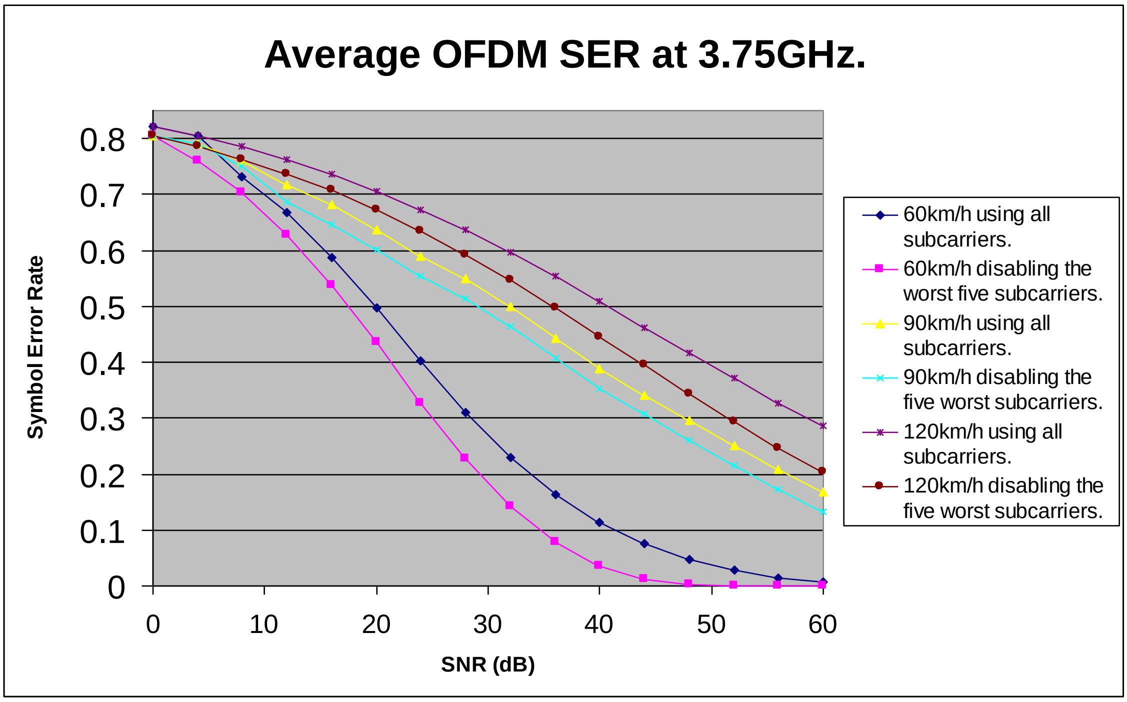


Figure (4-2) System performance at 3.75GHz.

The same procedure was repeated for the frequency of 3.75GHz, as figure (4-2) shows. Beside the previous notes concerning figure (4-1), a larger drift is clear in the curves, especially at higher speeds of 90km/h and 120km/h. For example, at speed 90km/h and for an accepted SER value of 0.2, a SNR value of 57dB is required when all subcarriers are used instead of only 45.5dB which was the value at 2.5GHz. A value of approximately 54dB is needed when the worst subcarriers were disabled. That gives an improvement of 3dB.

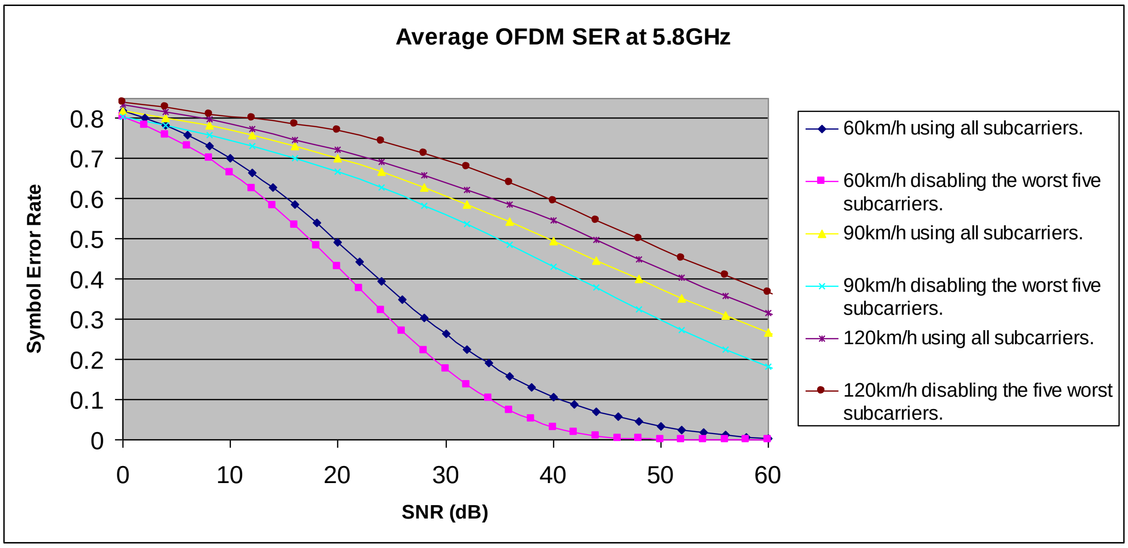


Figure (4-3) System performance at 5.8GHz

Figure (4-3) shows the OFDM system performance at 5.8GHz. More drift is clear in the curves, and larger SNRs are required to achieve a specific value of SER.

Each value of the SNR corresponds to a value of throughput/capacity according to the preceding capacity formulas. Each of these capacity values where calculated for two of the previous cases, for 60km/h at 2.5GHz, and for 120km/h at 5.8GHz. calculations where performed by equation (3-13). Then the capacity difference was plotted versus SNR, as shown in figure (4-4) a and b.

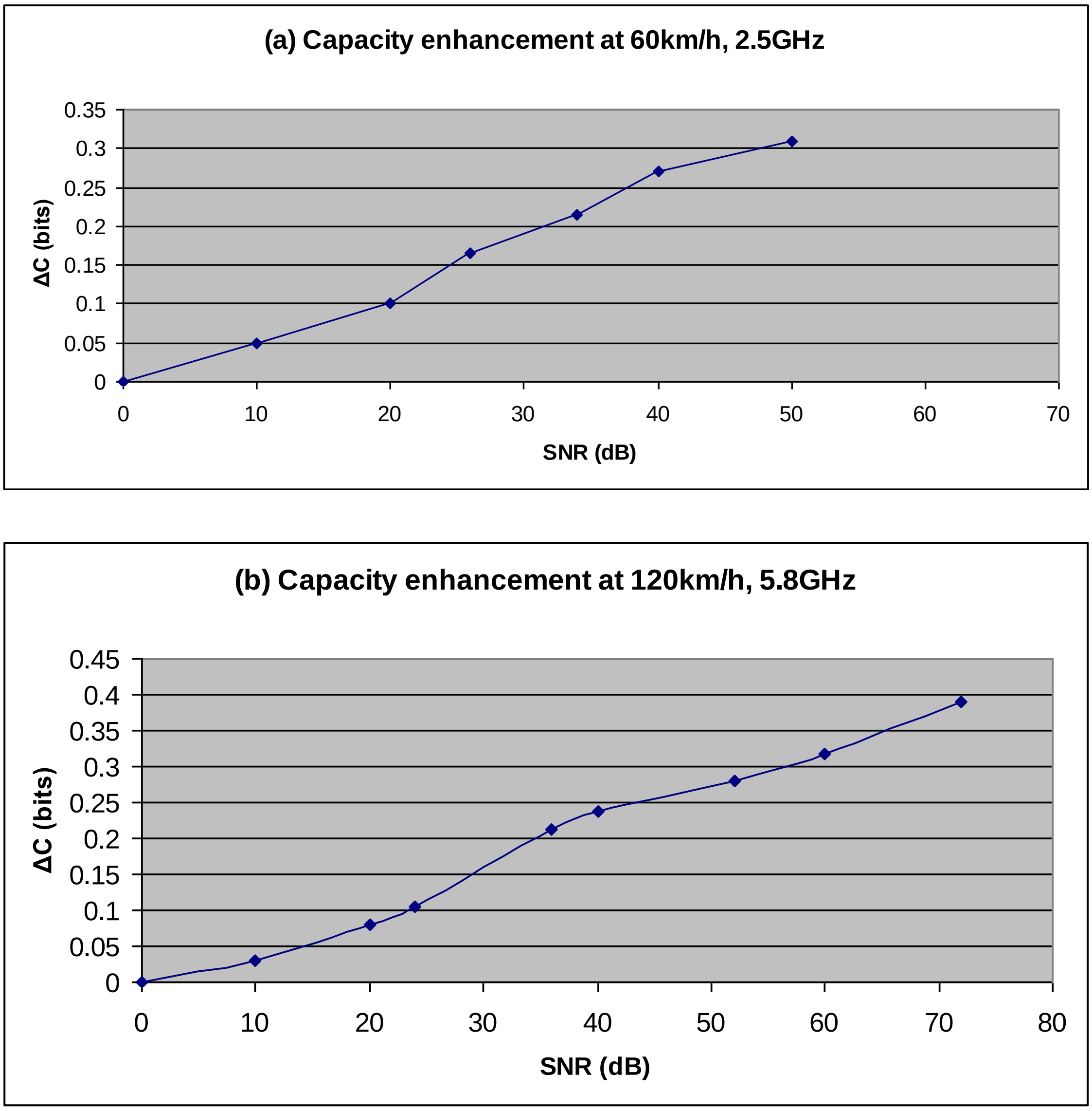


Figure (4-4) Capacity Enhancement

From figure (4-4) the following notes can be extracted:

- A general well noticed enhancement in the system's performance at all speeds and all frequencies. This is represented by the difference between the system's capacity while suffering from Doppler, and the same system's capacity with partial Doppler (the enhanced system).

- A linear relation between the capacity enhancement and the SNR at lower speeds and frequencies. This is obvious in figure (4-4)-a, where delta C is plotted for a speed of 60km/h at 2.5GHz carrier frequency.
- A linear relation between the capacity enhancement and the SNR at higher speeds and frequency carriers. This is clear in figure (4-4)-b, where delta C is plotted for a speed of 120m/h at ٥.٨GHz carrier frequency.
- For lower speeds and frequencies, the maximum capacity improvement reached approximately 3٥% at a SNR value of 50dB. Meanwhile a value of 5% enhancement was achieved for lower SNRs.
- At a higher speed and frequency, a maximum of also 33% capacity enhancement was achieved at a SNR value of 60dB.
- Higher values of capacity enhancement was obtained at lower frequencies, which might be due to higher Doppler frequency shifts at higher frequencies and speeds both at higher bit rates which make the improvement of the system capacity harder.

CONCLUSIONS AND RECOMMENDATIONS

The core idea in this research; that is to prove that for an OFDM/OFDMA system suffering from the Doppler Effect, the capacity /throughput can be increased by handing the users served by the worst subcarriers to a neighboring second or third cell (base station) according to the direction and distance of each user. This is applied when the users are away of the center of the serving cell. It must be done dynamically in a way similar to the way the AMC is applied. In this time the home base station could serve another five new users with good conditions on the mentioned subcarriers, in order to maintain the utilization of the system.

Now for users suffering from this phenomena, and passing near the center of the cell, those can be given a larger guard band at worst case. Accordingly that implies a waste of data which reduces the systems capacity in turn, but thankfully without QoS degradation as in the case of taking all subcarriers into consideration "fully Doppler".

It must be mentioned here that this simulation is done for one transmission path of the multipath system or MIMO system, or equivalently for a Single-input single-output (SISO) system. The same scenario is applicable for every single transmission path of the MIMO system and the overall capacity is the sum of the capacity on all paths, as discussed in chapter three. The MIMO system in turn, improves the capacity by diversity gain, as stated in chapter one.

The next step in this area would be putting new algorithms for the dynamic assignment of subcarriers for users taking into consideration these results, i.e. from Doppler Effects point of view.

The future work is going towards finding new methods for estimating the angle of arrival of the signals in order to determine the peaky angle, in order to take the

appropriate decision. In other words improving and applying adaptive angle estimation methods.

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APPENDIX

Matlab Code

The following lines contain the two Matlab files used to simulate the performance of the system under study. When run, the first program calls the second. The first program is edited each time to represent a different motion velocity and subcarrier frequency, in order to give a unique case.

```
%Matlab Program
% This Matlab code performs simulation of
% an OFDM system that employs QAM-16 baseband signaling
% a multipath channel with AWGN.
% Correct carrier synchronisation is assumed.
% 32 subcarriers are used with channel length of 6
% and cyclic prefix length of 5.
 clear;
%clf;
 L=1600000; % Total data symbols are 1 million
Lfr=L/32; % number of data frames
% Bit Error Rate for 16-QAM modulation using Gray modulation mapping
 N = L; % number of symbols
 M = 16;   % constellation size
k = log2(M); % bits per symbol
Spd=120;
Fcr=3.75;
Vlght=3;
DFSub=3200;
 DltaF=Spd*Fcr*10/(Vlght*3.6);
 % defining the real and imaginary PAM constellation
% for 16-QAM
alphaRe = [-(2*sqrt(M)/2-1):2:-1 1:2:2*sqrt(M)/2-1];
alphaIm = [-(2*sqrt(M)/2-1):2:-1 1:2:2*sqrt(M)/2-1];
    % symbol generation
    % ------------------
    ipBit = rand(1,N*k,1)>0.5; % random 1's and 0's
    ipBitReshape = reshape(ipBit,k,N).';
    bin2DecMatrix = ones(N,1)*(2.^[(k/2-1):-1:0]);  % conversion from
binary to decimal
  % real
    ipBitRe =  ipBitReshape(:,[1:k/2]);
    ipDecRe = sum(ipBitRe.*bin2DecMatrix,2);
    ipGrayDecRe = bitxor(ipDecRe,floor(ipDecRe/2));
    % imaginary
    ipBitIm =  ipBitReshape(:,[k/2+1:k]);
    ipDecIm = sum(ipBitIm.*bin2DecMatrix,2);
    ipGrayDecIm = bitxor(ipDecIm,floor(ipDecIm/2));
    % mapping the Gray coded symbols into constellation
    modRe = alphaRe(ipGrayDecRe+1);
    modIm = alphaIm(ipGrayDecIm+1);
    % complex constellation
    s_data = modRe + j*modIm;
 channel =[0.3 -0.5 0 1 0.2 -0.3]; % channel in t-domain
hf=fft(channel,32); % find the channel in f-domain
 p_data=reshape(s_data,32,Lfr); % S/P conversion
 p_td=ifft(p_data); % IFFT to convert to t-domain
p_cyc=[p_td(end-4:end,:); p_td]; %add cyclic prefix
s_cyc=reshape(p_cyc,37*Lfr,1); % P/S conversion
 Psig=10/32; % average channel input power
chsout=filter(channel,1,s_cyc); %generate channel output signal
 clear p_td p_cyc s_data s_cyc; % release some memory
```

```
 noiseq=((randn(37*Lfr,1) + j*randn(37*Lfr,1)));
 SEReq=[];
 for ii=1:31,
    SNR(ii)=ii-1; % SNR in dB
    Asig=sqrt(Psig*10^(-SNR(ii)/10))*norm(channel);
    x_out=chsout+Asig*noiseq; % Add noise
    x_para=reshape(x_out,37,Lfr); %S/P conversion
    x_disc=x_para(6:37,:); %discard tails
    [a b] = size(x_disc);
    Dplr=DltaF*cos((rand(a,b/2)-0.5)*pi)/(DFSub);
    Dplrx=[(cos(Dplr)+j*sin(Dplr)) ones(a,2*b/4)];
    x_disc=x_disc.*Dplrx;
   xhat_para=fft(x_disc); % FFT back to f-domain
   z_data=inv(diag(hf))*xhat_para; % f-domain equalizing
   % Compute the QAM decision after equalization
    deq=sign(real(z_data))+sign(real(z_data)-2)+sign(real(z_data)+2)+
j*(sign(imag(z_data))+sign(imag(z_data)-2)+sign(imag(z_data)+2));
    %Now compare against the original data to compute SER
    SEReq=[ SEReq sum(p_data~=deq,2)/Lfr ];
end
for ii=1:9,
    SNRa(ii)=2*ii-2;
    Q(ii)=3*0.5*erfc(sqrt((2*10^(SNRa(ii)*0.1)/5)/2));
    %Compute the Analytical BER
end
 %call another program to display OFDM Analysis
 ofdmAz
*******************************************************************
% MATLAB PROGRAM <ofdmAz.m>
% This program is used to analyze the OFDM subcarriers and their
% receiver outputs.
%Plot the subcarrier gains

figure(2);
stem(abs(hf));
xlabel('Subcarrier label');
title('Subchannel gain');

%Plot the subchannel constellation scattering after OFDM
figure(3);

subplot(221);
plot(z_data(1,1:800),'.') % subchannel 1 output
ylabel('Imaginary');
title('(a) Subchannel 1 output');
axis('square');

subplot(222);
plot(z_data(10,1:800),'.'); % subchannel 10 output
ylabel('Imaginary');
title('(b) Subchannel 10 output');
axis('square');

subplot(223)
plot(z_data(15,1:800),'.'); % subchannel 15 output
xlabel('Real');
ylabel('Imaginary');
title('(c) Subchannel 15 output');
axis('square');
```

```
subplot(224)
plot(z_data(:,1:800),'b.'); % mixed subchannel output
xlabel('Real');
ylabel('Imaginary');
title('(d) Mixed OFDM output');
axis('square');

 %Plot the average OFDM SER versus SER under "ideal channel"
%By disabling 5 poor subcarriers, average SER can be reduced.

Mat1=mean(SEReq);
Mat2=mean([SEReq(1:14,:);SEReq(20:32,:)]);

VV=[SNR;Mat1;Mat2]

figure(4);
figc=semilogy(SNRa,Q,'k-',SNR,mean(SEReq),'b-
o',SNR,mean([SEReq(1:14,:);SEReq(20:32,:)]),'b-s');
set(figc,'LineWidth',2);
legend('Ideal channel','Using all subcarriers','Disabling 5 worst
subcarriers')
title('Average OFDM SER');
axis([1 30 1.e-4 1]);
hold off;
xlabel('SNR (dB)');
ylabel('Symbol Error Rate (SER)');
```

أثرموقع و حركة المستخدم على سعة أنظمة التردد المقسمة عموديا ومتعددة الوصول.

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ملخـــــــــــــص

تظهر فائدة استخدام تقنية الإرسال متعدد الحوامل المتعامدة بتقسيم التردد عند ما تكون قناة الإتصال انتقائية التردد وبشكل كبير.ويعود السبب إلى إرسال المعلومات باسخدام حوامل فرعية متعامدة، بحيث تكون سرعة الرمز أقل من سرعة الرمز الطويل ذو الإرسال الطويل، مما يؤدي إلى توفير مناعة أعلى للنظام ضد خفوت الإشارة الناتج عن تعدد المسارات.

يعد الحفاظ على تعامد الحوامل الفرعية في هذا النظام أمرا حرجا، ويتأثرأداء النظام بشكل كبير في حال فقدان هذه الخاصية.هناك أسباب مختلفة لفقدان التعامد ما بين الحوامل، مثل الخطأ في التردد، أوالانزياح الزمني، ولعل أهمها هو انزياح التردد نتيجة ظاهرة دوبلر أو ما يسمى بإنزياح دوبلر. تنشأ هذه الظاهرة نتيجة حركة المستقبل أو المرسل أو كليهما معا، بحيث يحدث في مثل هذه الحالة ما يسمى بتداخل الحوامل، والذي يؤدي إلى تداخل المعلومات المحملة على حامل فرعي، مع المعلومات المحملة على الحوامل الفرعية المجاورة.

تناولت هذه الدراسة تأثير ظاهرة دوبلر على نظام التردد المقسم عموديا ومتعدد الوصول ومحاولة التقليل والحد من أثرها، وذلك عن طريق الأخذ بعين الإعتبار مقدار الانزياح في التردد الناتج عن هذه الظاهرة وتسليم المستخدمين الأكثر تأثرا بهذه الظاهرة إلى خلية اتصال أخرى مجاورة ليحل مكانهم مستخدمين آخرين بحالة أفضل بالنسبة لهذه الظاهرة.

وقد خلصت الدراسة إلى أنه باتباع هذه الطريقة، يمكن تحسين أداء النظام وبنسب تراوحت ما بين ٥% عند قيم الإشارة إلى الإزعاج المتدنية، و٣٥% عند قيم الإشارة إلى الإزعاج المرتفعة نسبيا.